IX. COMMUNICATION RESEARCH

A. MULTIPATH TRANSMISSION

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1. Speech and Music. Transatlantic Tests

Tests are in progress on 11.6575 Mc/sec. The interference from broadcast and code stations has been so severe that special regenerative band elimination filters have had to be made. They have resulted in an improvement of 30-40 db in the resultant signal-to-interference ratio. We now have a margin of 20-30 db between the minimum signal and maximum interference as averaged on a fast pen recorder. Pulse patterns indicate spreads of as much as 4 msec in transmission time over the various effective ionospheric paths. The few tests carried out, using peak deviations of $\pm$ 10 kc, compare favorably on intelligibility with commercial telephone circuits, but are not of good enough quality for broadcast relaying. The possibility of improvement by the use of diversity receptions is being considered.

J. Granlund, C. A. Stutt, L. B. Arguimbau

2. Television

The nature of the resultant signal produced by two-path transmission has been discussed in earlier reports. When the two signals are nearly equal in strength the frequency of the resultant signal is near the average frequency most of the time but is widely different for short periods. These "frequency spikes" occur at the difference frequency and are such that the average frequency of the resultant is exactly that of the larger signal (Interference in Frequency-Modulation Reception, Technical Report No. 42, p. 6, Research Laboratory of Electronics, M.I.T. January, 1949). In the case of sound transmission these spikes usually occur at a supersonic rate and can be averaged out. For picture transmission this is no longer the case and the spikes appear as a number of black or white lines following any sharp transition.

The effects of the spikes can be reduced by the use of de-emphasis, the video signal being pre-emphasized before transmission. The pre-emphasis of a video waveform gives rise to overshoots which must be removed with clipping circuits to prevent band-width limitations from being exceeded. The removal of these overshoots results in a deterioration of picture quality. The present investigations are directed towards producing the best compromise which is to be compared with pictures obtained using amplitude modulation.

R. D. Stuart, E. E. Manna, G. M. Rodgers

3. Simplified FM Receiver

The typical requirements of a commercial FM receiver are being studied. For this purpose the Massachusetts area has been mapped to find the normal and also the worst interference conditions met for FM broadcasts. This is being done as a practical example of the selectivity requirements that a receiver should meet. The receivers we have been using on the multipath project have differed from commercial models in many ways. In particular the linear i-f amplifiers have had less variation in transmission within the passband than is usual. This renders the selectivity problem more difficult. An effort is being made to meet both the flatness and selectivity problems without getting a prohibitively elaborate design.

It should be noted that this project is being carried on with the objective of narrowing the gap between the complex laboratory receiver we have been using and the home-type and communication-type receivers now commercially available. We are not trying to produce a completed commercial design but rather to study the applicability of the laboratory techniques to receivers that are within the reach of commercial practice.

R. A. Paananen

B. STATISTICAL THEORY OF COMMUNICATION

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1. Multichannel Analog Electronic Correlator

The design of the five-channel correlator has been nearly completed. Pulse distribution circuits and the circuits comprising the A sampling-pulse channel have been constructed. The remainder of the equipment is now being built. Various direct coupled integrators are being investigated in order to obtain a simple circuit that is acceptably drift-free and has a time constant of approximately 250 seconds.

Y. W. Lee, J. F. Reintjes, M. J. Levin

2. Analog Electronic Correlator for Second-Order Correlation

If $f_1(t)$ is a member function of a stationary random process, the second-order autocorrelation function is defined as

$$\phi_{111}(\tau_1, \tau_2) = \lim_{T\to\infty} \frac{1}{2T} \int_{-T}^{T} f_1(t)\, f_1(t+\tau_1)\, f_1(t+\tau_2)\, dt \quad . \tag{1}$$

This expression is a time average for ϕ_{111} and the equivalent ensemble average is

$$\phi_{111}(\tau_1, \tau_2) = \int_{-\infty}^{\infty}\int_{-\infty}^{\infty}\int_{-\infty}^{\infty} y_1 y_2 y_3\, p(y_1, y_2, y_3\,;\, \tau_1, \tau_2)\, dy_1 dy_2 dy_3 \tag{2}$$

where $p(y_1, y_2, y_3\,;\, \tau_1, \tau_2)$ is the second joint probability distribution density of the random process.

If at times t, $t + \tau_1$, and $t + \tau_1 + \tau_2$ the values of the ensemble are $a_1, a_2, a_3, \ldots$: $b_1, b_2, b_3, \ldots$; and $c_1, c_2, c_3, \ldots$ respectively, then Eq. 2 may be expressed as

$$\phi_{111}(\tau_1, \tau_2) = \lim_{N\to\infty} \frac{1}{N} \sum_{n=1}^{N} a_n b_n c_n \Big|_{\tau_1, \tau_2} \quad . \tag{3}$$

For large values of N, the approximate expression is

$$\phi_{111}(\tau_1, \tau_2) \cong \frac{1}{N}\sum_{n=1}^{N} a_n b_n c_n \Big|_{\tau_1, \tau_2} \qquad (4)$$

Equation 4 may be evaluated electronically by means of a device employing analog-computer techniques. The waveforms in Fig. IX-1 show the method of carrying out the computation. Sections of the wave in Fig. IX-1a, each of duration T, form an ensemble of random waves. During each T period, three samples of the signal amplitude are taken. Thus if the sampling pulses are separated by intervals τ_1 and τ_2 as indicated in Fig. IX-1b, the corresponding signal amplitudes at the instants of sampling are a_1, b_1, and c_1 for the first sampling period, a_2, b_2, and c_2 for the second sampling period, and so on (see Fig. IX-1a).

In order to obtain the product of the a, b, and c samples as required by Eq. 4, a rectangular pulse of amplitude proportional to a and width proportional to b is first generated. The area under this pulse is therefore proportional to the product ab (see Fig. IX-1c). A second pulse is then generated so that its amplitude is proportional to the product ab and its width is proportional to c (see Fig. IX-1d). The area under this pulse is the product abc.

The areas under the abc-product pulses resulting from N sets of samples may be summed by applying the product pulses to an integrator. The integrator output voltage after N sets of samples is then proportional to the second-order autocorrelation function of the signal for the particular delay intervals τ_1 and τ_2. A three-dimensional family of the correlation functions may be obtained by repeating the above steps for various values of τ_2, with τ_1 held constant at different values.

Figure IX-2 shows a block diagram of the circuits required to carry out the computations. It may be noted that the unit is similar to the single-channel analog electronic correlator now being used in the laboratory to compute first-order correlation functions. (See J. F. Reintjes: An Electronic Analog Correlator, Technical Report No. 189, Research Laboratory of Electronics, M.I.T., to be published). The importtant difference is that for the second-order computation, a third channel (the C channel in Fig. IX-2) is necessary in order to obtain three samples of the input signal during each sampling period.

Y. W. Lee, J. F. Reintjes

3. Cathode Ray Tube Display of Correlator Output

In order to improve on the usual graphic-recorder display of the correlator output, display on a cathode ray tube was tried. All the experiments were conducted on the analog electronic correlator (J. F. Reintjes: An Analog Electronic Correlator, Technical Report No. 189, Research Laboratory of Electronics, M.I.T., to be

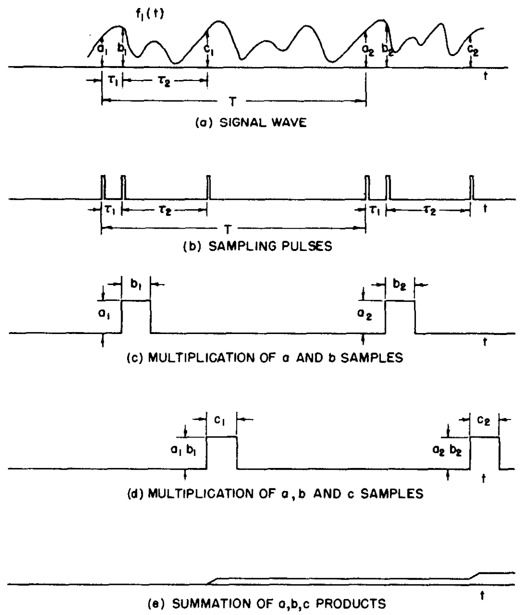

Fig. IX-1

Computation of second-order correlation functions by method of discrete sampling.

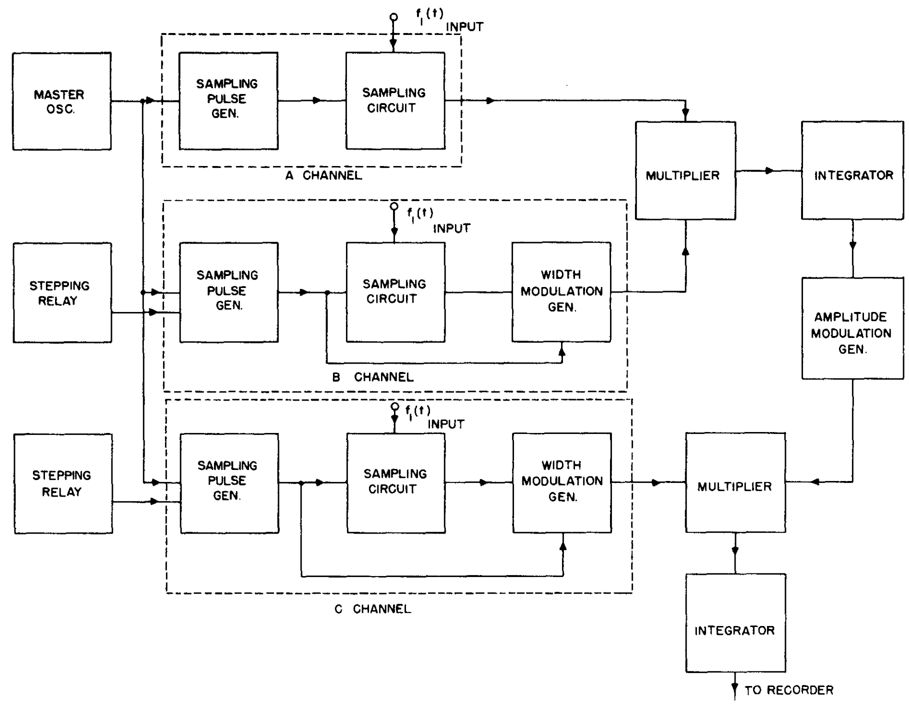

Fig. IX-2

Block diagram of an analog computer for evaluating second-order correlation functions.

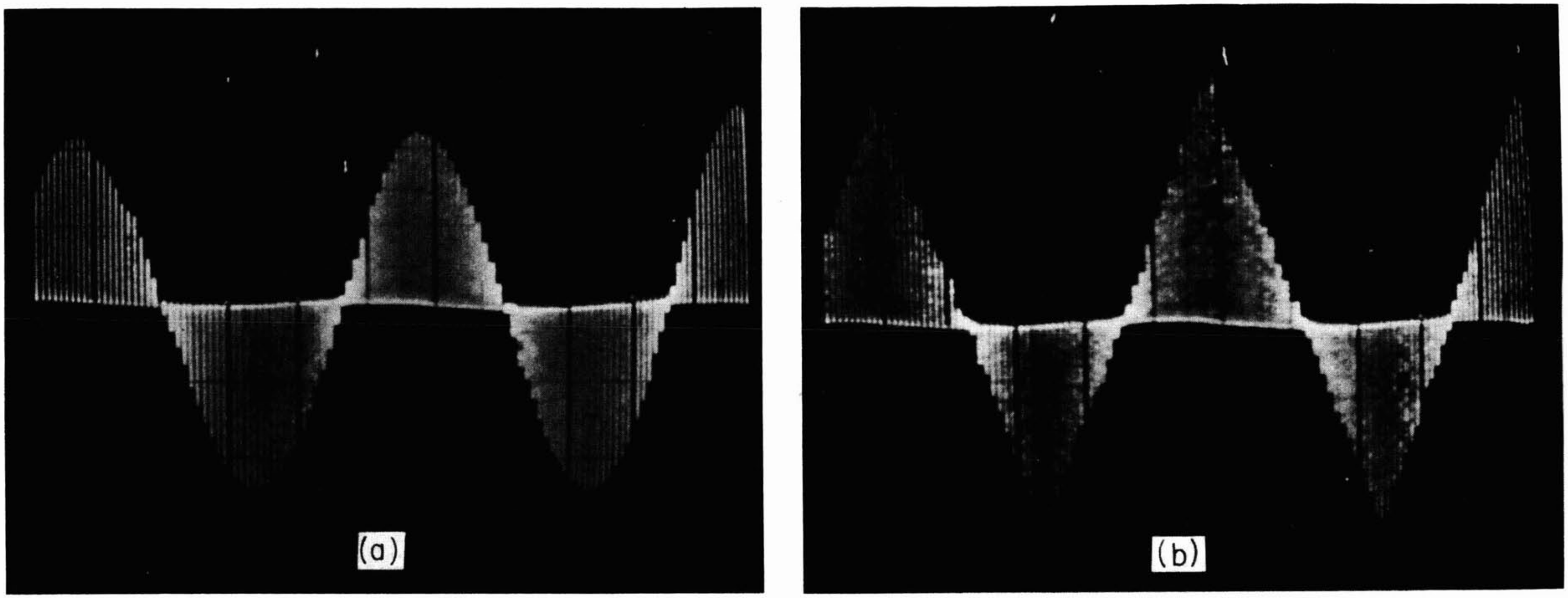

Fig. IX-3

Enlargements of time exposure pictures (Polaroid Land camera) of the autocorrelation functions of (a) a sine wave and (b) a square wave. $\omega_o = 4$ kc/sec; $\Delta\tau = 5$ μsec; 8000 samples/point.

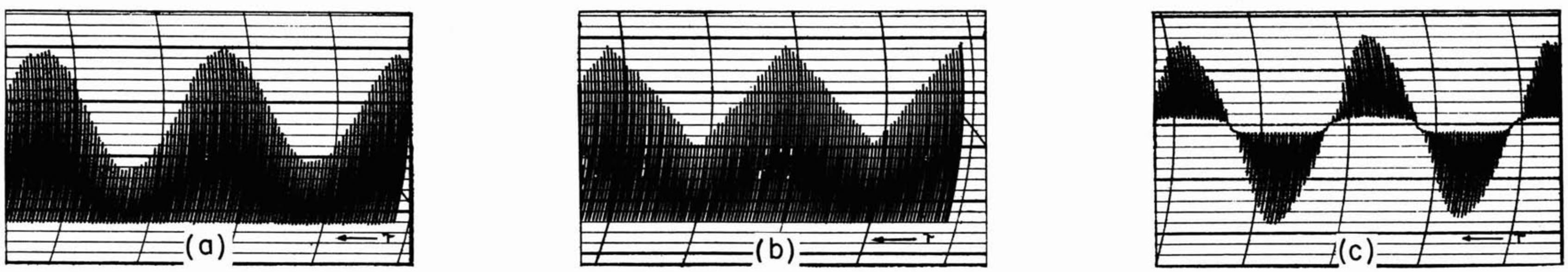

Fig. IX-4

Graphic records (Esterline-Angus recorder) of the autocorrelation functions of (a) a sine wave with bias added, (b) a square wave with bias added, (c) a sine wave with no bias added. $\omega_o = 4$ kc/sec; $\Delta\tau = 5$ μsec; 8000 samples/point.

published), of which the performance was being checked.

A horizontal sweep for the cathode ray tube was obtained from the stepping voltage generated within the correlator to control the trigger delay of the phantastron. The correlator output was d-c coupled to the vertical deflection plates. Figure IX-3 shows time exposure photographs taken with a Polaroid Land camera from the screen of a Tektronix Mod. 512 oscilloscope. A comparison of these correlation curves with the Esterline-Angus records in Fig. IX-4 is of interest.

1. A presentation in linear coordinates has been achieved with the cathode ray tube display.

2. On the cathode ray tube the display abscissa and the actual delay (τ) are controlled in synchronism, while on the graphic recorder the time base is driven completely independently of τ. The latter system is thus liable to errors if the stepping voltage or the drive speed varies irregularly. For example, the long τ step taken by the correlator every 10 points cannot be detected on the graphic records while it is immediately noticeable on the photographs.

3. Mechanical friction, inertia effects, pointer backlash and inking errors inherent to recorders have no counterpart in the cathode ray tube display method. The electron beam deflects faithfully in either direction. Sensitivity is increased since no bias is necessary.

4. The use of a Polaroid Land camera permits a picture to be developed within one minute of the end of a correlation run which lasts about 18 minutes. Resolution on the enlargements of the original picture is excellent.

5. Careful study of the photographs discloses the presence of switching transients between steps and of variations in the speed of writing. An effort is being made to perfect the machine performance on the basis of this information.

J. F. Reintjes, J. J. Bussgang, M. Coufleau

4. Crosscorrelation Functions Relating to Amplitude Clippers

Some interesting results concerning crosscorrelation functions of input and output $\phi_{12}(\tau)$ of a symmetrical two-level clipper (no delay) have been computed in terms of the input autocorrelation function $\phi_{11}(\tau)$.

In the case of a sinusoidal input

$$\phi_{12}(\tau) = \phi_{21}(\tau) = \frac{2\sqrt{2}}{\pi}\sqrt{\phi_{11}(0)\ \phi_{22}(0)}\ \frac{\phi_{11}(\tau)}{\phi_{11}(0)} \quad . \tag{1}$$

In the case of gaussian noise input

$$\phi_{12}(\tau) = \phi_{21}(\tau) = \sqrt{\frac{2}{\pi}}\sqrt{\phi_{11}(0)\ \phi_{22}(0)}\ \frac{\phi_{11}(\tau)}{\phi_{11}(0)} \quad . \tag{2}$$

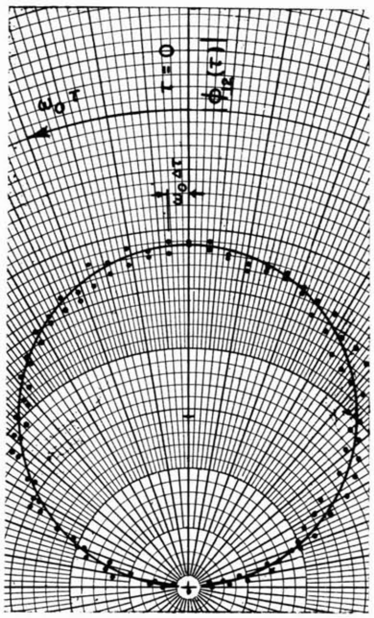

Fig. IX-5
Polar plot of one period of the crosscorrelation function $\phi_{12}(\tau)$ of a sine input and the clipper output. $\Delta\tau = 5$ μsec; $\omega_o = 2000$ cps; 8000 samples/point.

The above theoretical results have been confirmed experimentally within the accuracy of the electronic analog correlator (see above).

In evaluating $\phi_{12}(0)$ for a sine wave input to the clipper, polar diagrams were used and the radius of the circle interpolated (see Fig. IX-5). In this way we used all the points of the curve instead of using just the peak points and taking their average.

J. B. Wiesner, J. J. Bussgang

5. Synthesis of Speech from Short Time Autocorrelation Function

An experiment has been performed to establish the possibility of speech synthesis from a short time autocorrelation function. Available for this purpose was the M.I.T. short time speech correlator (1, 2) which provides an analysis of the speech signal into a thirteen-point representation of the autocorrelation function. Since no machine has yet been built to perform the corresponding synthesis operation directly in the time domain the following indirect method was employed. The cosine transformation of the short time correlation function was generated giving the short time power spectrum (3), which was then combined using the synthesizer section of an eight-channel Bell Telephone Laboratories vocoder (4) to form speech. The cosine transformation, which in this case amounts to a linear transformation of the correlator output signals with constant coefficients, was derived by means of a resistance matrix or weighting circuit. Each resistance R_{mn} in the 8 by 13 array was chosen to correspond to a particular delay τ_m in the corresponding correlator channel, and the particular center frequency

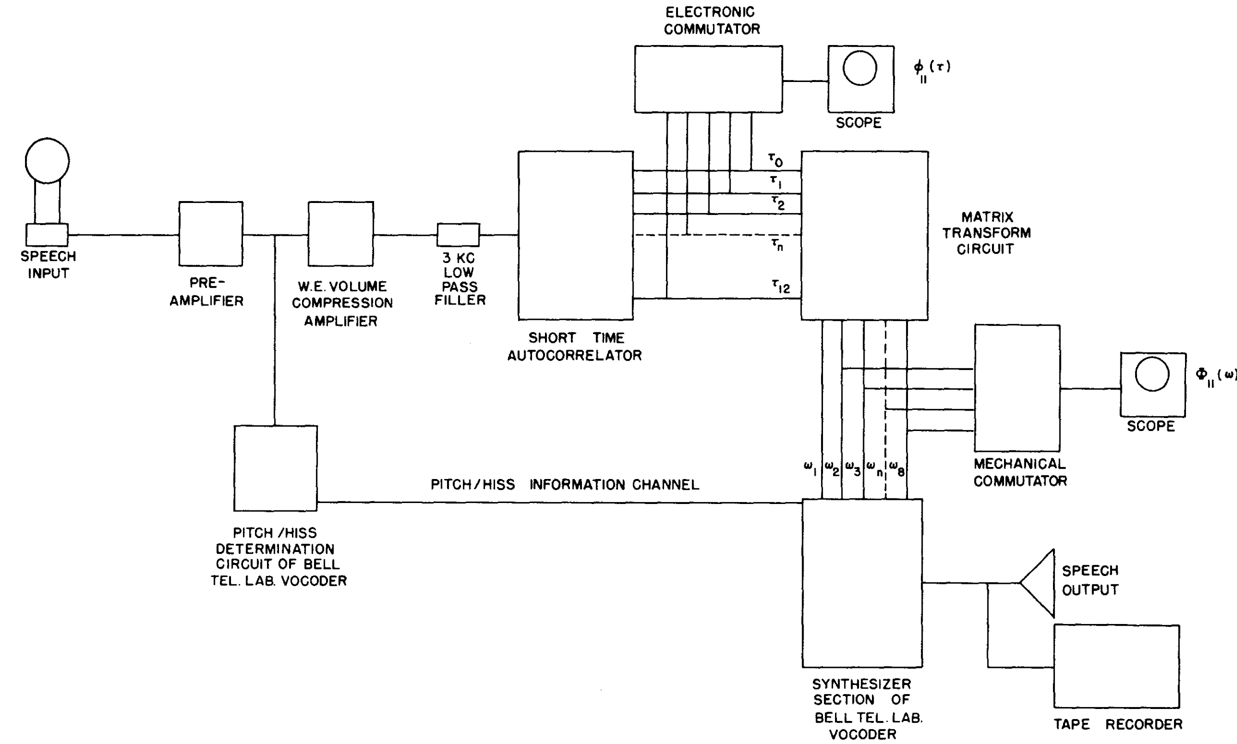

Fig. IX-6

Block diagram of experiment for synthesizing speech from short time autocorrelation function.

ω_n of a vocoder channel as follows: $R_{mn} = |1/(\cos \tau_m \omega_n)|$. (Negative coefficients were effected by means of phase inverters in each of the correlator channels and the resistances were connected to the appropriate polarity signal.) A block diagram of the whole setup is shown in Fig. IX-6; a further complication necessary for the experiment was the pitch/hiss channel input to the synthesizer which was supplied from the corresponding section of the vocoder analyzer. This additional information about the speech signal was found to be essential for producing intelligible speech output. Electronic and mechanical commutators were used to monitor the correlation function and frequency spectrum signals, respectively, at the input and output of the matrix, and to facilitate alignment of the system. The speech output of the synthesizer was recorded on a magnetic tape and records of preliminary tests are on file.

Two tests of the system performance were made: First we tested the frequency response of the correlator and matrix transform circuit for sine-wave input to the correlator shown in Fig. IX-7. Here the output voltage of several of the frequency channels is plotted against the frequency. It will be noted that this curve is a sin ω/ω curve displaced by the amount of the center frequency; it is acutally just the Fourier

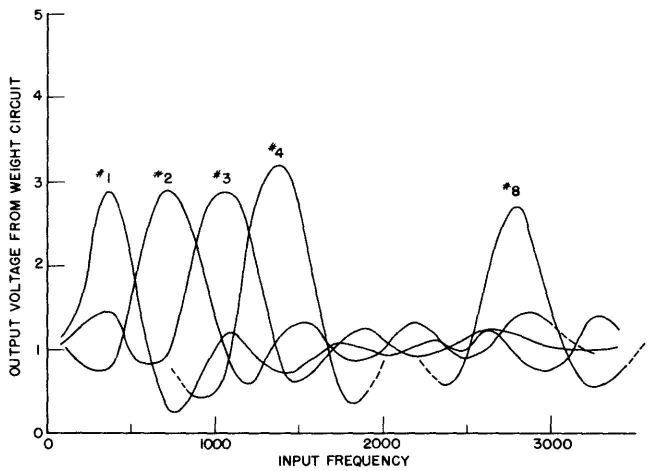

Fig. IX-7

Frequency response curves for autocorrelator and matrix transform circuit.

transform of the boxcar-type weighting function in the time domain implicit in the use of only thirteen points to represent the correlation function. These response curves can be modified by giving different weight to the different delay channels; this would permit reduction of the spurious negative responses at the sacrifice of sharpness of the resonance peaks; however, no measurements of the effect of different weighting functions on the over-all speech intelligibility of the system have been made. The second was an over-all test with speech passed through the whole system; the word articulation score was not measured but was estimated at 30 percent, enough to enable understanding of simple sentences and other highly redundant text. There was some indication in these tests that the response of the integrating filters in the correlator output channels, approximately 20-cycle cutoff, was too slow to follow rapid speech articulation; further tests should be made with faster integrating circuits and other refinements.

The main purpose of this experiment was to show that the essential information required to convey speech intelligibility is contained in the short time correlation function, in particular as derived by the present model speech correlator, with the possible necessary addition of an extra channel to convey information concerning the excitation function, i.e. whether the sound was voiced or unvoiced. The test described above was imperfect in realizing the full information content of the correlator output signals: (a) information is certainly lost in the linear transformation from the thirteen correlator signals to the eight channels of the synthesizer and (b) the frequency response characteristics of the synthesizer did not match exactly those of the correlator matrix combination; indeed it would require some sort of different weighting function to give a realizable, i.e. nonnegative, frequency response for the correlator matrix combination.

It is believed possible to build a much better synthesizer than the one described above, with all thirteen channels using direct synthesis in the time domain and with an exact match of frequency response characteristics which would perhaps give a much higher articulation score; it would show much better the true possibilities of the correlation-function representation of the speech wave (5).

Such a system would be the exact analog of the vocoder system which uses frequency band analysis, and might prove to be more practical to instrument, using timing and switching circuits in place of frequency filtering.

B. Howland, B. A. Basore, R. M. Fano, J. B. Wiesner

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2. Quarterly Progress Report, Research Laboratory of Electronics, M.I.T. p. 68, January 15, 1950

3. R. M. Fano: J. Acous. Soc. Am. 22, No. 5, 546, 1950

4. H. Dudley: J. Acous. Soc. Am. 11, 169, 1939

5. Quarterly Progress Report, Research Laboratory of Electronics, M.I.T. p. 70, January 15, 1950

6. Information Theory

a. Energy limitations in information theory

An attempt has been made to reconsider some fundamental physical facts in the light of the information theory. Taking into account only the basic sources of noise some physical consequences were obtained from the fundamental theorem of the information theory (1, 2).

Starting from Nyquist's theorem and the definition of the channel capacity, we have found that an amount of energy of at least $kT \log_e 2$ must be spent in order to transmit one bit of information.

Reasoning on Szilard's one-molecule engine (3) has shown that given one bit of information about the system an amount of energy $kT \log_e 2$ can be extracted from the heat reservoir at temperature T which is assumed to be surrounding the engine. It has further shown that $kT \log_e 2$ is the maximum amount of energy that can be obtained from one bit of information.

The above results are valid only in the range $h\nu \ll kT$.

We have analyzed a specific system operating in the range $h\nu \gg kT$ and found that the limit is essentially $kT \log_e 2$ ergs/bit.

This work will be published in the near future.

C. A. Desoer, R. M. Fano

References

1. C. E. Shannon: Mathematical Theory of Communications, p. 39, University of Illinois Press, 1949

2. R. M. Fano: Transmission of Information II, Technical Report No 149, p. 14, Research Laboratory of Electronics, M.I.T. Feb. 1950

3. L. Szilard: Zeit. f. Physik 53, 840, 1929

b. Transmission of information through channels in cascade

This work is a first step in the generalization of information theory towards a general theory of communication through an arbitrary network.

It was first assumed that there is no storage decoding and recoding between two consecutive channels in order to approximate the situation occurring in practice.

Discrete and continuous channels have been considered. Theoretical developments were carried out to enable study of the variation of the channel capacity with respect to the number of channels. Numerical results have been obtained for binary channels.

A pulse code modulation system has been compared to a continuous type of transmission; the great difference in performance tends to show that even the rough form of coding of the former can considerably improve the behavior of the system.

C. A. Desoer, R. M. Fano

C. HUMAN COMMUNICATION SYSTEMS

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P. F. Thorlakson

The work of this group will be reported at a later date.

D. PARALLEL CHAIN AMPLIFIER

The parallel chain amplifier is a very wide band one that overcomes the basic gain bandwidth limitation of conventional amplifiers. The method is to divide the band into parts small enough so that each may be amplified by a conventional chain. These chains are then paralleled to produce an amplifier of arbitrary gain and bandwidth.

As was mentioned in the Quarterly Progress Report, July 15, 1951, a frequency-scaled model has been built to demonstrate the theory involved. Work has continued in the direction of building a full-scale model. The 6AK5 pentode was chosen and its input and output admittance measured as a function of frequency in a variety of circuits. The circuit which gave the best results was used to build a four-tube, stagger-tuned chain. A minimum of 16-db gain over the band 130-260 Mc was obtained.

The reason for the difficulty in amplifying in the uhf region is a loading produced by the transit time effect and the lead inductance of the tubes. Attention was focused on the grid circuit admittance, where this loading is most pronounced. Measurements were made with a G. R. admittance meter and associated equipment, including the AN/APR-1 as a null detector. This setup gave repeatable, consistent results and was simple to operate. The grid terminals were connected to the meter by a coaxial cable and hence a Smith chart was needed to correct the readings. The results were obtained as a plot on the Smith chart and proved useful as a comparative device between different connections.

Some of the circuits used are shown in Fig. IX-8. The main difference lies in the connections made to the two cathode leads. In Fig. IX-8a, one cathode lead is used for the grid return and the other for the plate and screen return. This connection minimizes the inductance common to the grid and plate circuits. The Smith chart plot of the input admittance has a loop which goes deep into the heavy conductance area of the chart and appears unusable much above 200 Mc.

In the circuit of Fig. IX-8b, the plate and screen return is made to ground and the second cathode lead is unused. This circuit resulted in a plot which had a cusp at about 280 Mc which protruded into the heavier conductance region. The protrusion was not so severe as to make the tube unusable in this region (that is, to have a maximum gain less than unity). The plot above the cusp showed good results up to about 350 Mc.

The circuit of Fig. IX-8c is a combination of the previous two circuits in that the plate and screen return is made to both the cathode and ground. This connection gave a smooth plot with no loops or cusps and appeared useful to 300 Mc.

The circuit of Fig. IX-8d differs from Fig. IX-8c in that an interstage network of the type to be used has been inserted in the plate circuit. This interstage consists of the three coils shown plus the interelectrode capacitances which shunt the end coils (the series capacitor is used for blocking) and has two resonant frequencies. When these

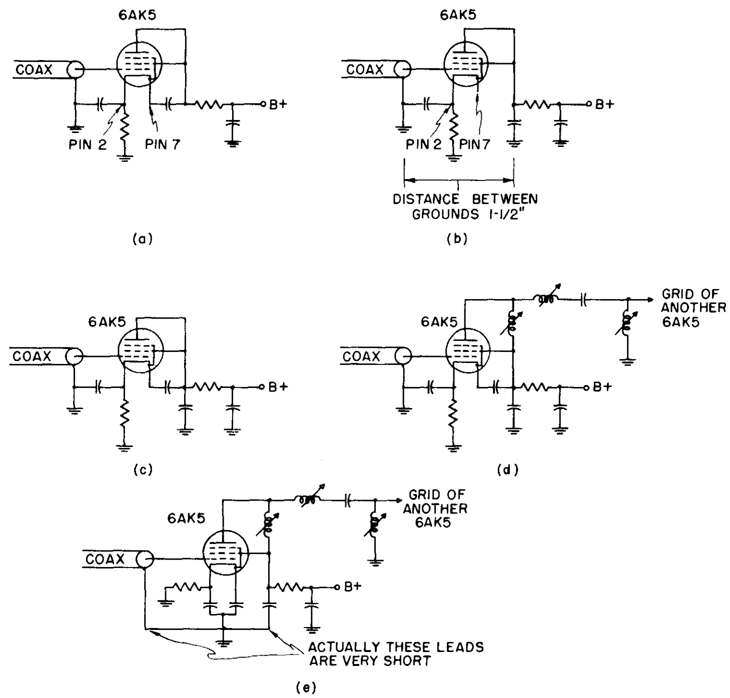

Fig. IX-8

Circuit connections. All resistors are 200 ohms; all capacitors are 0.001 µf. B+ is 122 volts.

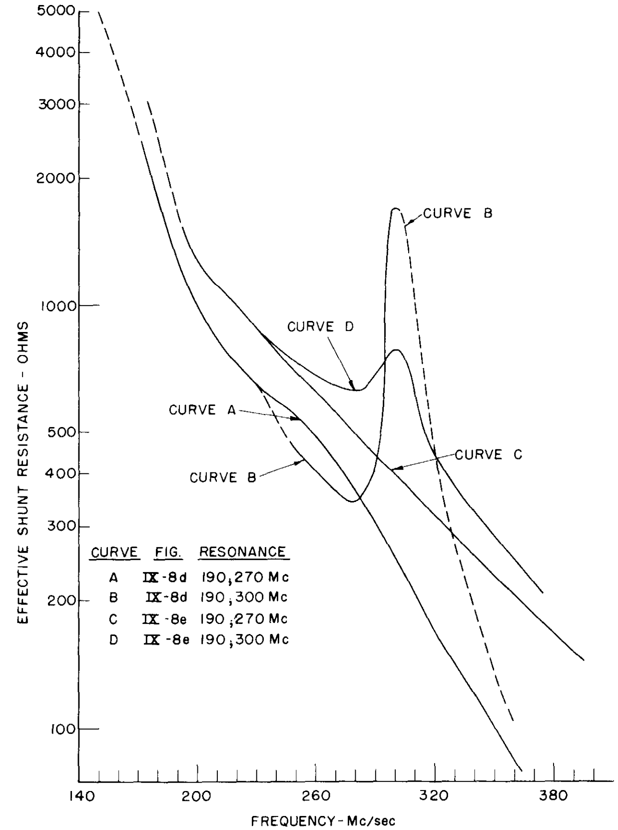

Fig. IX-9

Effective input shunt resistance - 6AK5.

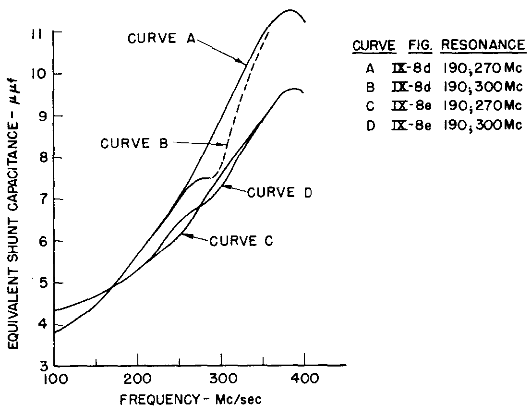

Fig. IX-10

Effective input capacitance - 6AK5.

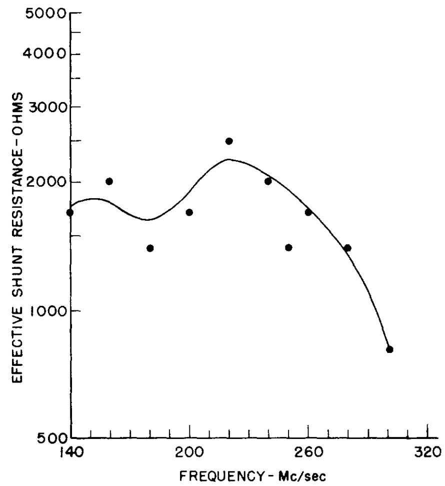

Fig. IX-11

Effective output shunt resistance - 6AK5.

two resonant frequencies are set at 190 and 270 Mc the admittance plot is essentially the same as that for Fig. IX-8c. However, for resonances at 190 and 300 Mc a pronounced hump was noticed around 300 Mc in the direction of smaller conductance. This change gives hope of using the tube at 300 Mc and above. The data obtained for these two conditions are plotted in Figs. IX-9 and IX-10 as equivalent shunt resistance and equivalent shunt capacitance, respectively, as functions of frequency.

The connections of Fig. IX-8e are a result of discussing the problem with an engineer of the RCA Application Engineering Department. He mentioned that minimizing the cathode lead inductance was more important than minimizing the lead inductance common to the grid and plate circuits. The circuit of Fig. IX-8e minimizes the cathode lead inductance by effectively paralleling the two leads, at the expense of greatly increasing the common inductance. His advice proved sound and results better than those from Fig. IX-8d were obtained. These are plotted in Figs. IX-9 and IX-10.

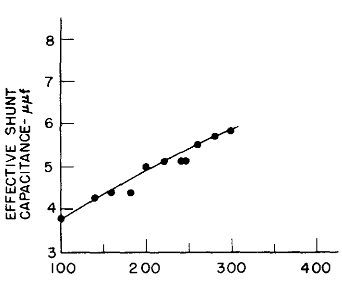

Fig. IX-12

Effective output shunt capacitance - 6AK5.

The output admittance (between plate and ground) for the connection of Fig. IX-8c is indicated in the plots in Figs. IX-11 and IX-12.

The chain which was built was simply a succession of four circuits like that of Fig. IX-8e terminated in another 6AK5 which served a dual role as a load and an isolation for the measuring equipment. The coils were all slug-tuned, with the slugs isolated from ground to avoid excessive stray capacitances. Gain measurements were

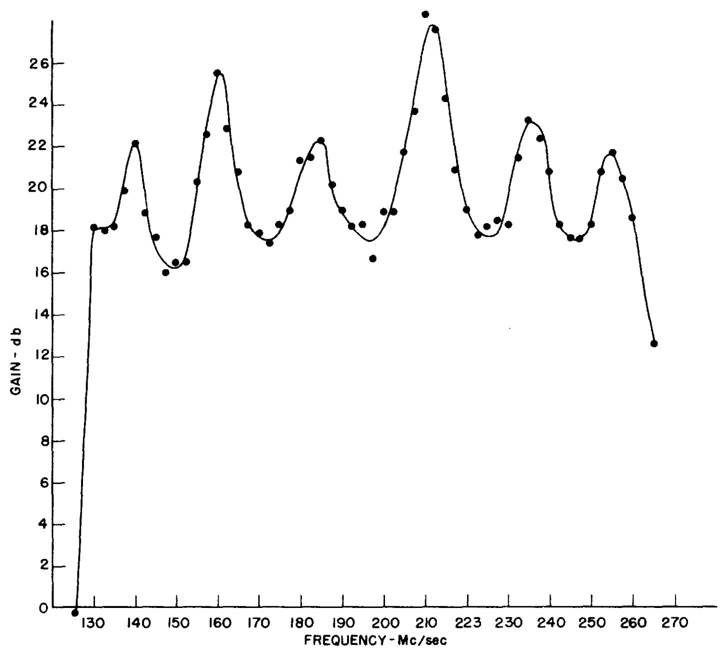

Fig. IX-13
Amplitude response of chain.

made by the substitution method using a G. R. 1021 AV signal generator and the APR-1 as a detector. An accuracy of better than 1 db is expected. The amplitude characteristic is plotted in Fig. IX-13. No attempt to smooth off the peaks was made, although this could be done by resistive padding with a sacrifice in gain. When considering these results, it should be noted that additional stages would be tuned to fill in the valleys so that it is reasonable to take the gain of these four stages as its average, say 20 db. The gain per stage then is 5 db, assuming at least several stages. This compares favorably with the maximum theoretical gain using two terminal pair interstages, without any shunt conductance, which is 6.1 db per stage. This is based on the following capacitances:

Tube Input	6 $\mu\mu f$	Includes socket capacitances
Tube Output	5	(see Figs. IX-10 and IX-12).
Stray	4	
	15 $\mu\mu f$	

It is hoped that in the near future another chain to cover the range 0-130 Mc will be built and the two combined. A penalty of the order of 10 db is expected for paralleling. Our consolation is that this penalty need be paid only once, no matter how many tubes are used.

R. K. Bennett, J. G. Linvill

E. EXPERIMENTAL APPROXIMATION AND NETWORK ALIGNMENT

In the Quarterly Progress Report, July 15, 1951, a method was described to facilitate the alignment of a network to bring its response to a standard signal into correspondence with a desired response. The method involves application of the fact that small changes in the parameters of the network result in changes of the response which are linearly related to the amount of element change. The technique suggested employs the generation of the functions which represent the change in response with small unit changes in each of the elements. A further experimental step generates a family of normal orthogonal functions which are linearly dependent upon these functions. The orthogonal functions are then used essentially to analyze a misaligned network to prescribe the changes in it to bring its response to that desired.

A simple laboratory test has been run which checked the practicability of the alignment technique with affirmative results. A block diagram of the experimental setup is shown in Fig. IX-14. The networks used were simple low-pass filters, one constructed of fixed elements and two with adjustable elements. The network configuration and approximate element values are given in Fig. IX-15. The network with fixed elements is the standard network and its response to the square wave is the standard or desired response. The adjustable networks are used for multiple purposes. If in one of them all elements but one are adjusted to the standard values, with the one element a unit different, the connected amplifier of Fig. IX-14 receives a voltage which is the change in response corresponding to a unit change in that element. By setting a number of the elements of the adjustable network in the proper manner (see Quarterly Progress Report, July 15, 1951) one can cause the amplifier to receive a voltage which is one of the orthogonal functions mentioned. Finally, one of the networks can simulate a network being aligned, while the companion adjustable network is successively adjusted to present

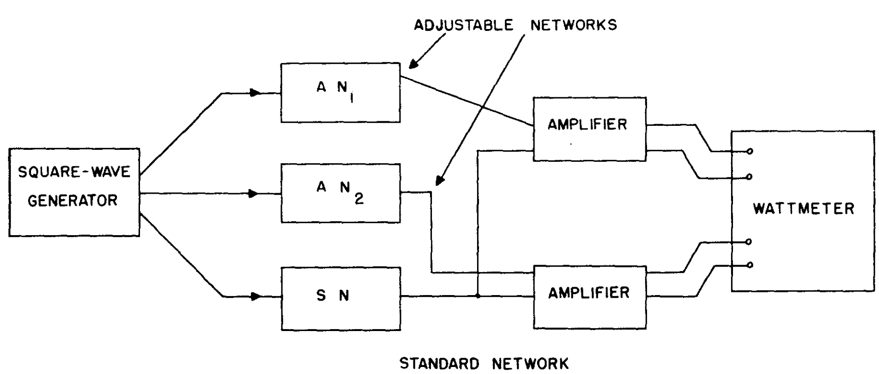

Fig. IX-14
Block diagram of test setup.

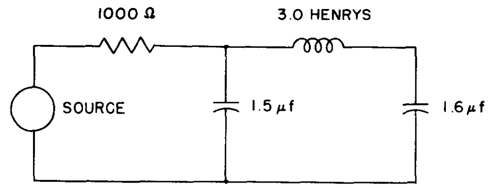

Fig. IX-15
Configuration and approximate element values of networks used.

the family of orthogonal functions to the amplifier. The corresponding succession of wattmeter readings shows the changes which should be made to bring the misaligned network into alignment. The wattmeter is a simple form of instrument which shows the average product of two variables fed to it. The amplifiers of Fig. IX-14 are used as differencing circuits. The square wave has a repetition rate of 30 cps; this frequency is chosen to make the wattmeter most effective as an averaging multiplier.

The experimental results will be covered fully in a forthcoming technical report. It is clear from results observed with this simple example that the method is practically workable, and that a wattmeter is a very effective averaging multiplier in this frequency range. It is proposed to make further study of the technique in practical instances at higher frequency ranges.

As indicated earlier, the same technique of adjustment can be applied in the approximation problem solution by successive approximations. The question of designing suitable model networks for such a device is being investigated by W. I. Wells.

J. G. Linvill

F. NEW METHODS OF NETWORK SYNTHESIS

Technical Report No. 201 has been prepared and is scheduled for publication. This report presents synthesis procedures that realize practical RLC and RC networks. The RLC networks are practical in that they contain no mutual inductance and no perfect coils, i.e. every inductance has an associated series resistance. New techniques employed in the procedures are first discussed in detail and then applied to various synthesis problems. Included among the procedures for synthesizing RLC networks are those for realizing unbalanced structures and lattices whose arms possess identical poles. Reduction of the lattices to unbalanced forms is considered, and it is shown that if real transformers are allowed, i.e. transformers with winding resistance, magnetizing inductance and a coupling coefficient smaller than one, then the lattice realization for a transfer admittance is always reducible.

Unbalanced networks are realized directly by the RC synthesis procedure. The number of elements required is smaller than that required by the Guillemin method of RC synthesis.

Finally, two new synthesis procedures are presented for realizing a Darlington network without any ideal or unity coupled transformers, where a Darlington network is considered to be composed of lossless elements plus only one resistance. In one of the two procedures the single resistance appears not as a termination but within the network.

L. Weinberg, E. A. Guillemin

G. TRANSIENT PROBLEMS

1. Basic Existence Problems

The report on this work will be resumed in the Quarterly Progress Report of January 15, 1952.

2. Network Synthesis for Prescribed Transient Behavior

This problem has been completed and will be presented in Technical Report No. 209.

W. H. Kautz